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TITLE: SIGNAL PROCESSING METHOD AND DEVICE FOR A SPREAD SPECTRUM RADIO COMMUNICATION RECEIVER

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SIGNAL PROCESSING METHOD AND DEVICE FOR A SPREAD SPECTRUM RADIO COMMUNICATION RECEIVER

BACKGROUND OF THE INVENTION

The present invention relates to systems for radio communication with mobiles. It lies within the receivers used in the fixed or mobile stations of such systems and operating coherent demodulation of spread spectrum signals.

Coherent demodulation requires various parameters representing the propagation channel between the transmitter and the receiver. Some of these parameters vary relatively slowly and can be estimated by statistical probing methods. Such is the case for example for the delays assigned to the multiple propagation paths in the conventional rake receiver. The delays specific to the various paths can be updated at fairly low frequency, for example of the order of a second. On the other hand, other parameters have abrupt variations, on the scale of the duration of an information symbol, due to the fading phenomenon. Such is the case in particular for the instantaneous amplitudes of reception of the symbols along the propagation paths taken into consideration, which are required for coherent demodulation. These instantaneous amplitudes are complex amplitudes, manifesting the attenuation and the phase shift undergone at each instant along the paths.

In general, these complex amplitudes are estimated from symbols known a priori, or pilot symbols, interspersed among the transmitted information symbols so as to

allow coherent demodulation. This mode of estimation does not permit allowance for the fast channel variations between the patterns of pilot symbols.

An object of the present invention is to improve the consideration of fast fading in coherent demodulation schemes.

SUMMARY OF THE INVENTION

The invention thus proposes a method of processing a digital signal at the output of a filter matched to a spreading code in a spread spectrum radio communication receiver, the digital signal comprising successive blocks each corresponding to a sequence of symbols sent by a transmitter, each sequence comprising at least one symbol known a priori and information symbols. The method comprises an estimation of statistical parameters representing a channel having at least one propagation path between the transmitter and the receiver; and a processing of each block of the digital signal to estimate instantaneous amplitudes of reception of the symbols of the corresponding sequence sent by the transmitter. According to the invention, the processing of a block comprises the estimation of a group of at least one information symbol of the sequence by optimization of a criterion defined by the digital signal of said block, the estimated statistical parameters, at least one symbol of the sequence which is known a priori and each information symbol of said group; and the estimation of said instantaneous amplitudes as a function of the digital signal of said block, of the estimated statistical parameters, of the symbols of the sequence which are known a priori and of

the estimated symbols of said group.

The method performs a joint optimization of some at least of the information symbols with the instantaneous amplitudes of reception of the symbols, thereby making it possible to improve the reliability of the estimations.

The method is applicable when the signals are transmitted on two parallel sub-channels between the transmitter and the receiver, for example two quadrature sub-channels only one of which comprises the pilot symbols. This is the case for the uplink, from the mobile terminals to the base stations, in the third-generation cellular systems of UMTS type ("Universal Mobile Telecommunications System"). The information symbols estimated jointly with the instantaneous amplitudes can then all be transmitted on the same sub-channel as the pilot symbols, these amplitudes subsequently being used to perform coherent demodulation on the other sub-channel.

The method is also applicable by performing the joint estimation only on certain of the information symbols time-division multiplexed with the pilot symbols, for example in the case of the downlink, from the base stations to the mobile terminals, in UMTS systems.

Another aspect of the present invention relates to a method of estimating the speed of movement of a mobile radio communication station based on a digital signal produced by a filter matched to a spreading code in a spread spectrum radio communication receiver, the digital signal comprising successive blocks each corresponding to a sequence of symbols sent by a

transmitter, each sequence comprising at least one symbol known a priori, said mobile station comprising one of said transmitter and receiver. This method comprises

- storing a table of vectors for a collection of values of speed of movement of the mobile station, the table having, for each speed value, at least one entry containing an autocorrelation vector of instantaneous amplitudes of reception of symbols sent by the transmitter, precalculated according to a propagation model;
- estimating the autocorrelation vector of the instantaneous amplitudes on the basis of the portions of the blocks corresponding to the symbols known a priori;
- selecting an entry of the table of vectors, containing the precalculated autocorrelation vector closest to the estimated autocorrelation vector; and
- estimating the speed of movement of the mobile station on the basis of the selected entry.

It is thus possible to achieve reliable estimations of the speed of the mobile station, which may in particular be used in a signal processing method as defined above.

The invention also proposes signal processing devices tailored to the implementation of the above methods.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is a schematic diagram of an exemplary radio

communication receiver incorporating the invention.

Figure 2 is a schematic diagram of a channel analysis and symbols estimation unit of the receiver of Figure 1.

DESCRIPTION OF PREFERRED EMBODIMENTS

The invention is described below within the framework of a spread spectrum radio communication system using a code-division multiple access technique (CDMA), of which UMTS is an example. A channel of such a system on a carrier frequency is defined by a spreading code composed of discrete samples called "chips", having real values (± 1) or complex values (± 1 ± j), which follow one another at a chip rate F_c (F_c = 3.84 Mchip/s in the case of UMTS).

We consider the reception of a CDMA radio signal block along a multipath channel having additive white noise, the block resulting from a sequence of N symbols which is produced by a transmitter. The symbols are real valued (± 1) or complex valued (± 1 ± j). The duration $1/F_s$ of a symbol on a channel is a multiple of the duration of the chip, the ratio of the two being the spreading factor $Q = F_c/F_s$ of the channel. In the example of UMTS, a block can correspond to a timeslot of a 10 ms radio frame (i.e. 666 µs of signal since a frame comprises 15 timeslots), the spreading factor Q being a power of 2 lying between 4 and 256 with Q.N = 2560 chips.

Moreover, L denotes the number of propagation paths allowed for by the receiver, and W the length of the impulse response of the channel, expressed as a number

of chips. The propagation profile of the channel is defined by a set of eigenvectors v_i and of associated eigenvalues λ_i for $0 \le i \le L$. Each eigenvector v_i of dimension W, is a waveform associated with an echo in the impulse response of the channel. In a traditional "rake" receiver, each eigenvector v_i can have just one nonzero component, corresponding to a propagation delay allocated to a finger of the receiver. More generally, these vectors v_i can have several nonzero components.

Each sequence of N symbols comprises a number p_0 of symbols which are known a priori, or pilot symbols. We are interested here in the reception of the portion of the block corresponding to a collection of p symbols of the sequence, including at least one information symbol unknown a priori. It is assumed that these p symbols comprise the p_0 pilot symbols ($p_0 < p \le N$). They could however comprise just some (at least one) of the pilot symbols. The signal observed for the estimation of the parameters of the channel, composed of Q.p complex samples $y_0, y_1, \ldots, y_{Q.p-1}$, may be written:

$$Y = M.V.B(b).P.A + N' \tag{1}$$

where:

- $A = (A_0, A_1 \ldots, A_{L-1})^T$ (with $A_i = \left(a_i^0, a_i^1, \ldots, a_i^{N-1}\right)$ and $(.)^T$ designating the transposition operation) is a column vector with L.N components a_i^n corresponding to the complex amplitudes (instantaneous fading realizations) for the various symbols, indexed by n, of the block and the various paths, indexed by i:

- $P = \begin{pmatrix} \Pi & 0 & \cdots & 0 \\ 0 & \Pi & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & \Pi \end{pmatrix}$ is a puncturing matrix of size $L.p \times L.N$, in which a puncturing pattern Π of size $p \times N$ is repeated L times along the diagonal, the q-th row of the pattern Π $(1 \leq q \leq p)$ being composed of N-1 times the value 0 and once the value 1 at the position corresponding to the q-th symbol of the collection of p symbols (P is the identity matrix of size L.N when p = N);

- $b = (b_0, b_1, \dots b_{p-1})^T$ is a vector of p components which are equal to the p symbols b_0, b_1, … b_{p-1} of said collection;

- $B(b) = \begin{pmatrix} B'(b) & 0 & \cdots & 0 \\ 0 & B'(b) & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & B'(b) \end{pmatrix}$, with $B'(b) = \begin{pmatrix} b_0 & 0 & \cdots & 0 \\ 0 & b_1 & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & b_{p-1} \end{pmatrix}$ is a diagonal matrix of size $L.p \times L.p$ where the diagonal sub-matrix $B'(b)$ is repeated L times along the diagonal;

- V is a matrix of size $W.p \times L.p$ containing the eigenvectors v_i of the channel, which are assumed constant over the length of the block, i.e. $V = (V_0, V_1, \dots, V_{L-1})$ where V_i is a matrix of size $W.p \times p$ where the column vector v_i is present p times: $V_i = \begin{pmatrix} v_i & 0 & \cdots & 0 \\ 0 & v_i & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & v_i \end{pmatrix}$;

- M is a convolution matrix for convolving with the channel spreading code, of size $Q.p \times W.p$, whereby

the code portions corresponding to the p symbols observed have been concatenated;

- N' is a column vector of size Q.p composed of samples of noise assumed to be additive and gaussian; and
- $Y = (y_0, y_1, \ldots, y_{Q.p-1})^T$

To estimate the $p - p_0$ unknown symbols and the L.N components of the vector A jointly, we seek to maximize the conditional probability $pr(A,b \mid Y)$, which is proportional to $pr(Y \mid A,b).pr(A,b)$. Since A and b are independent and all the sequences of bits are assumed equiprobable, we have to maximize $pr(Y \mid A,b).pr(A)$, which is equivalent to minimizing the criterion:

$$\|Y - M.V.B(b).P.A\|^2 + N_0.A^H.K^{-1}.A \qquad (2)$$

where N_0 is the variance of the noise, and K the autocorrelation matrix of the fading $K = E(A.A^H)$, where E(.) designates the mathematical expectation and $(.)^H$ the conjugate transpose.

By differentiating the criterion (2) with respect to the components of A, we can express A as a function of b and reintroduce the expression obtained in (2), this leading to searching for the vector $\hat{b}$ which maximizes the criterion:

$$Z^H.B(b).P.\left(P^H.P + N_0.K^{-1}\right)^{-1}.P^H.B(b).Z \qquad (3)$$

with:
$$Z = V^H.M^H.Y \qquad (4)$$

We can subsequently obtain the estimate $\hat{A}$ of the vector A as a function of that $\hat{b}$ of the vector b:

$$\hat{A} = \left(P^H.P + N_0.K^{-1}\right)^{-1}.P^H.B(\hat{b}).Z \qquad (5)$$

By means of this estimate Â, the remaining information symbols can be subjected to coherent demodulation in a "rake" receiver of conventional type. These remaining symbols can be those which have not already been estimated on the basis of the current block (in the case where $p < N$).

If several blocks are transmitted in parallel, for example on two quadrature sub-channels, the estimate Â is used to demodulate a block transmitted in parallel.

Figure 1 shows a CDMA receiver of the latter type, which processes blocks transmitted in parallel on two quadrature sub-channels (I and Q). This receiver can belong to a base station ("node B") of a UMTS type network in FDD mode ("Frequency Division Duplex"). The I sub-channel (real part of the complex baseband signal) transports only data bits, while the Q sub-channel (imaginary part) transports N control bits with a spreading factor $Q = 256$. These N control bits include p_0 pilot bits and $N - p_0$ bits carrying control information unknown a priori to the receiver. For a precise description of these uplink channels, reference may be made to the technical specification 3G TS 25.211, version 3.3.0, "Physical Channels and Mapping of Transport Channels onto Physical Channels (FDD) (Release 1999)", published in June 2000 by the 3GPP ("3rd Generation Partnership Project"), Section 5.2.1. The number p_0 of pilot bits lies between 3 and 8, and the other control bits include $N_{TPC} = 1$ or 2 bits carrying transmission power control commands, $N_{TFCI} = 0$, 2, 3 or 4 bits indicating a transport format combination used on the channel and $N_{FBI} = 0$, 1 or 2 feedback information bits.

The receiver illustrated by Figure 1 comprises a radio stage 1 which performs the analogue processing required on the radio signal picked up by the antenna 2. The radio stage 1 delivers a complex analogue signal whose real and imaginary parts are digitized by the analogue/digital converters 3 on respective I and Q processing sub-channels. On each sub-channels, a filter 4 matched to the shaping of the pulses by the transmitter produces a digital signal at the chip rate of the spreading codes.

On the I sub-channel, this signal is subjected to a matched filter 5 corresponding to the spreading code c_I assigned to the data bits of the channel. The resulting signal is processed by a conventional rake receiver 6 which delivers estimates $\hat{d}$ of the transmitted data bits.

On the Q sub-channel, another matched filter 7, operating with the spreading code c_Q with spreading factor Q = 256 transforms each digital signal block Y from the filter 4 into a block X submitted to a channel analysis and control bits estimation unit 8. The unit 8 supplies the rake receiver 6 with the eigenvectors v_i relating to the L propagation paths considered, as well as the estimated instantaneous complex amplitudes $\hat{a}_i^n$ $(0 \le i < L,\ 0 \le n < N)$ corresponding to the N = 10 bits transmitted on the Q sub-channel.

With the notation of model (1) and with p = N, the signal block X of size W.N delivered by the matched filter 7 can be written in the form of a column vector $X = M^H.Y$.

A schematic diagram of the unit 8 is represented in

Figure 2. The units 10 surrounded by a broken line serve to calculate statistical parameters representing the propagation channel between the transmitter and the receiver, namely the eigenvectors v_i supplied to the rake receiver 6, the estimate N_0 of the variance of the noise and the parameters characterizing the autocorrelation matrix $K = E(A.A^H)$.

If it is assumed that the realizations of the fading are independent random variables for two different paths, then we can write:

$$K = \begin{pmatrix} K_0 & 0 & \cdots & 0 \\ 0 & K_1 & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \cdots & 0 & K_{L-1} \end{pmatrix} \quad (6)$$

where K_i is an autocorrelation matrix of size $N \times N$ relating to path i. It is further assumed that these autocorrelation matrices K_i are proportional, i.e. $K_i = \lambda_i . \overline{K}$, λ_i being the eigenvalue corresponding to the eigenvector v_i of the path, and $\overline{K}$ being normalized for a unit energy. The eigenvector decomposition of the matrix $\overline{K}$ can be written:

$$\overline{K} = \sum_{k=0}^{F-1} \mu_k . f_k . f_k^H \quad (7)$$

where F is the number of eigenvectors to be taken into consideration which (like the eigenvectors and eigenvalues in question) depends on the speed of movement of the mobile terminal. The normalization of $\overline{K}$ implies that $\sum_{k=0}^{F-1} \mu_k = N$.

The eigenelements μ_k, f_k can be calculated by estimating the matrix K and by extracting them via an

appropriate algorithm. Another solution, requiring less computation power, consists in selecting these eigenelements as a function of an estimate of the speed of the mobile station.

The correlation matrix K is thus characterized by the quantities λ_i, μ_k and f_k for $0 \le i < L$ and $0 \le k < F$. These elements are supplied to a joint estimation module 20, which minimizes the criterion (3), which is equivalent to maximizing the criterion:

$$\sum_{i=0}^{L-1}\sum_{k=0}^{F-1}\frac{\lambda_i \cdot \mu_k}{\lambda_i \cdot \mu_k + N_0}\left|\mathbf{f}_k^H.\mathbf{B}'(b).\mathbf{Z}_i\right|^2 \qquad (8)$$

where $Z_i = V_i^H.X$. The module 20 determines the estimate $\hat{b}$ by maximizing criterion (8), then it obtains the estimates $\hat{a}_i^n$ by applying formula (5), which may also be written, in the example considered:

$$\hat{A}_i = \sum_{k=0}^{F-1}\frac{\lambda_i \cdot \mu_k}{\lambda_i \cdot \mu_k + N_0}\left(\mathbf{f}_k^H.\mathbf{B}'(\hat{b}).\mathbf{Z}_i\right)\mathbf{f}_k^T \qquad (9)$$

It is noted that the coefficients used in equation (9) have already been calculated when maximizing the criterion (8).

The convolution product, followed by a projection, $Z = V^H.M^H.Y = V^H.X = \left(Z_0^T, Z_1^T, \ldots, Z_{L-1}^T\right)^T$, is calculated by a module 21 over the entire length of the block so as to be processed by the joint estimation module 20.

In order to minimize (3) or maximize (8), the module 20 can undertake an exhaustive calculation of the values of the criterion according to the various possible values of the unknown control bits, and retain the set of values which yields the optimal value. To do this,

it can take advantage of the redundancy which may exist between certain of the bits of the control sub-channel. For example, when $N_{TPC} = 2$, it is in fact the same power control bit which is transmitted twice, thereby reducing the number of combinations of bits which have to be tested.

The units 10 determine the eigenvectors and eigenvalues v_i, λ_i in a conventional manner from the portions of the signal blocks corresponding to the pilot bits. The module 11 extracts these portions of the successive blocks, and the module 12 estimates, over these portions, the mathematical expectation of the matrix $X.X^H$. This may be performed by a calculation of a mean over around 100 blocks. The eigenelements v_i, λ_i are then calculated by the module 13 by diagonalization of the matrix $E(X.X^H)$, the L eigenvectors retained v_i being those for which the eigenvalues λ_i have the largest moduli.

By projecting the signal portions corresponding to the pilot bits onto the vectors v_i, the module 14 obtains the instantaneous amplitudes a_i^n relating to the p_0 pilot bits $(0 \leq n < p_0)$, as well as a residual value corresponding to a noise sample. The mean energy of these samples is evaluated by a module 15 in order to obtain the estimate of the parameter N_0. Moreover, a module 16 estimates, over an averaging window which may also be of the order of 100 blocks, the mathematical expectations of the quantities $a_i^n . a_i^{n+m*}$ which form the components $\hat{\gamma}_m = E\left(a_i^n . a_i^{n+m*}\right)$ of an autocorrelation vector $\hat{\gamma}$ of the instantaneous amplitudes $(0 \leq m < p_0)$.

The estimates N_0 and $\hat{\gamma}$ are supplied to a module 17 which estimates the eigenelements μ_k, f_k as well as the speed of movement of the mobile terminal. This module 17 cooperates with a memory 18 wherein is recorded a table T of autocorrelation vectors of eigenelements.

This table T contains autocorrelation vectors γ and sets of vectors of the eigenvalues f_k, μ_k for one or more radio propagation models and for several values of speed of movement of the mobile terminal.

By way of example, two types of propagation models, indexed by an integer m, may be taken into consideration, namely a Rayleigh channel and a Rice channel.

For each model m, and for various values of speed v, it is possible to calculate in advance the components of the autocorrelation vector, denoted $\gamma(m,v)$, of the instantaneous amplitudes for a noise level assumed zero, as well as the eigenelements f_k, μ_k of the matrix $\overline{K}$. An entry $T(m,v)$ respectively containing the vector $\gamma(m,v)$, the number $F(m,v)$ of eigenelements taken into consideration, and the eigenelements in question $f_k(m,v)$, $\mu_k(m,v)$ for $0 \leq k < F(m,v)$ is then recorded in the table T. This table T is calculated once and for all and recorded in the memory 18.

Subsequently, when the module 17 receives, for example every 100 blocks, estimates γ and N_0, it can select from the table 18 the autocorrelation vector which corresponds best to that which was estimated, while taking into account the presence of the noise on the channel. To do this, the module 17 can perform a

minimization in the least squares sense, i.e. find the model m and the tabulated speed v which minimize the quantity $\|\hat{\gamma} - N_0.\delta_0 - \gamma(m,v)\|^2$, where $\delta_0 = (1, 0, 0, \ldots, 0)^T$. The eigenvectors and eigenvalues f_k, μ_k which are located in the selected entry of the table T may then be supplied to the joint estimation module 20.

The minimization performed by the module 17 furthermore makes it possible to obtain an estimate of the speed of movement of the mobile terminal. This is the speed corresponding to the entry selected from the table T. This estimate v can be supplied to various other processing units of the radio communication system, making it possible to tailor the behavior of the system to the speed of the mobile terminals.

It has been found that the speed estimates obtained by this procedure were more reliable than those obtained by the conventional linear estimators.